A SYNCHRONOUSLY TUNED 60 MC I. F. AMPLIFIER WITH A. G. C.

By Richard Lane

A 5 stage transistor I. F. amplifier is described, having 70 db gain, 20 mc bandwidth and 40 db of A. G. C. using fixed neutralization and 2N916 transistors.

Previously there have been two main difficulties with the use of transistors in A. G. C. bandpass amplifiers, namely the dependence of the response on the parameters of individual transistors and the variation of these parameters and therefore the amplifier response with the application of A. G. C.

A design method is given to minimize the above effects. Due to the fact that for frequencies much higher than f_β the current gain falls at approximately 6 db per octave, the apparent center frequency of the amplifier is lower than the resonant frequency of the interstages. Expressions are derived in the appendix for the required interstage resonant frequency and the cascading shrinkage factor.

The designer of a vacuum tube bandpass amplifier has the choice of two arrangements of simple interstage, namely synchronous and stagger tuning to meet a given bandwidth requirement. The high conductance of transistors, however, limits gain/bandwidth trading to a stage bandwidth of $> \frac{G_T}{B_T} f_o$ mc/s, where

G_T = Total interstage conductance referred to transistor collector.
B_T = Total interstage susceptance referred to transistor collector.
f_o = Resonant frequency of interstage.

A synchronously tuned amplifier is simple to align and has a better pulse response for a given 3 db bandwidth than an amplifier with steeper response skirts.

For the above reasons synchronous tuning was chosen for this amplifier. Fixed neutralization is possible with these transistors because of the small spread in the already small y_{12_e}. In the absence of a good equivalent circuit for the "mesa" transistor both parallel and series R.C. neutralizing were evaluated by means of a sweep generator and it appeared that the parallel network gives broader band neutralizing than the series.

A. G. C. is obtained by controlling the emitter current of the second, third and fourth stages.

The first stage is fixed biased in order to preserve a good noise figure at all signal levels and the output stage in order to accomodate a large collector current swing without limiting.

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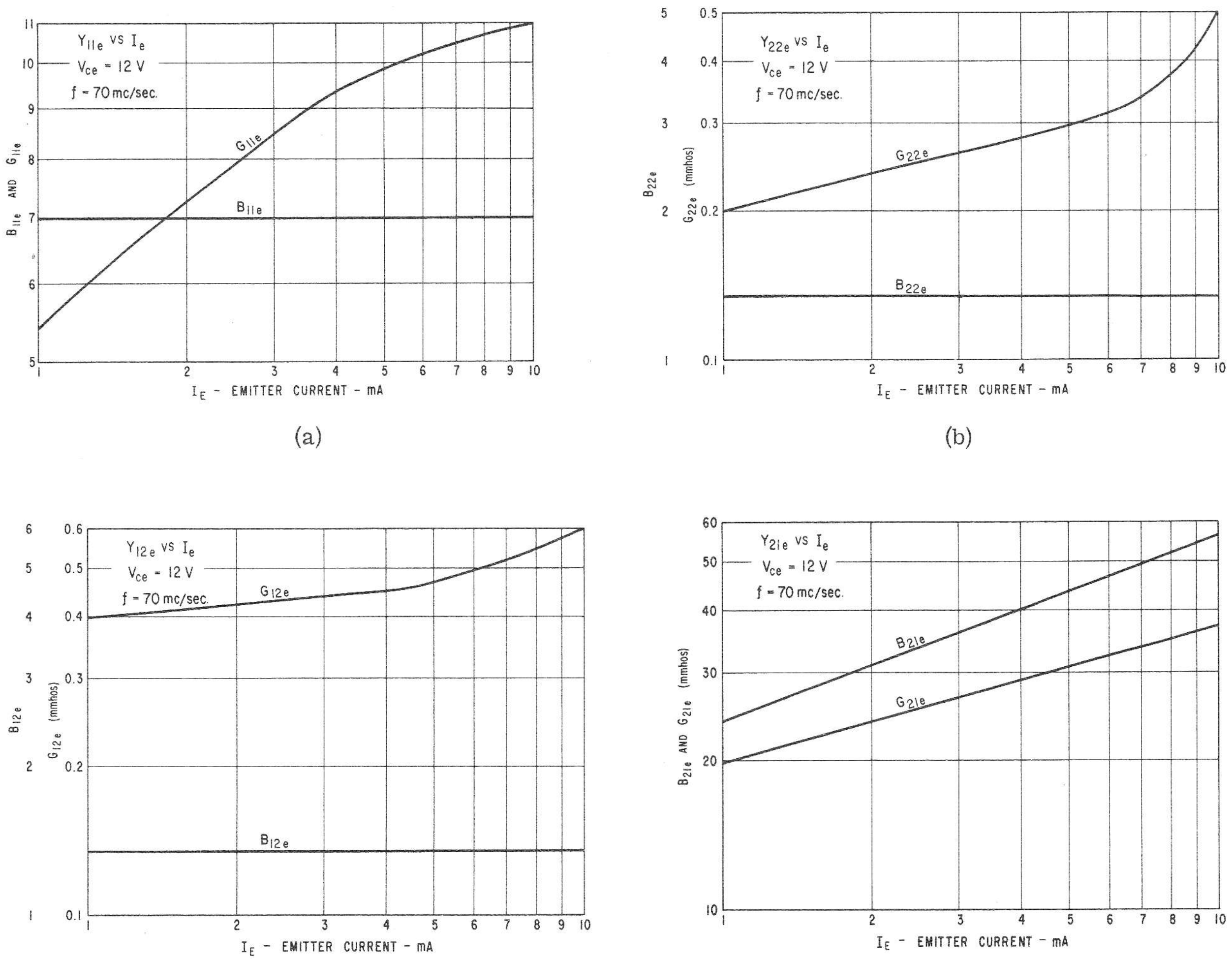

(a)

(b)

(c)

(d)

Fig. 1. Transistor y Parameters versus Emitter Current.

It is seen from the graphs of the transistor y parameters vs emitter current (Fig. 1) that when the emitter current is reduced the interstage susceptance changes only due to y_{12}, the feedback transfer admittance; but the conductance decreases considerably due to input and output conductance changes. This has the effect of increasing the interstage Q which while not changing f_o, the resonant frequency, causes f_{max}, the frequency of maximum gain, to increase according to the following relationship.

$$f_{max} = f_o \sqrt{1 - \frac{1}{2Q^2}}$$

This appears graphically in Fig. 2. A tangent of +6 db per octave is drawn to normalized single tuned response curves for various Q's, the point of tangency representing $\frac{\omega_{max}}{\omega_o}$ or $\frac{f_{max}}{f_o}$ is seen to approach unity as Q tends to ∞ .

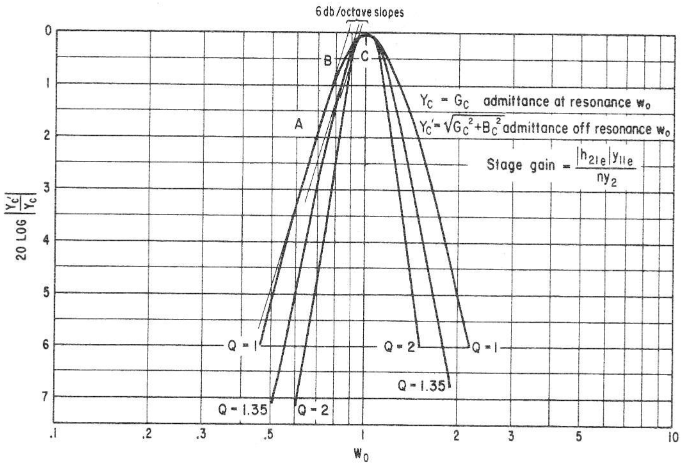

Fig. 2. Interstage Responses for 3 values of Q.

This frequency change resulting from the bandwidth narrowing could be corrected if a source of variable conductance were available. An FD 100 silicon diode has an admittance versus current relationship as follows:

$$y_{diode} = K_1 + K_2 (G + j\frac{G}{5}) I_d \text{ mmhos} \qquad \text{ref. (1)}$$

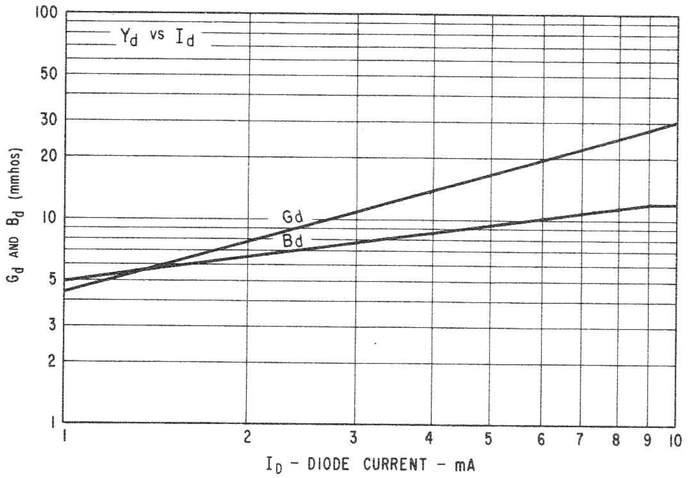

Fig. 3. Diode Admittance versus Current at 70 mc.

where

K_1 is $\ll y_c$ the admittance referred to the interstage primary.

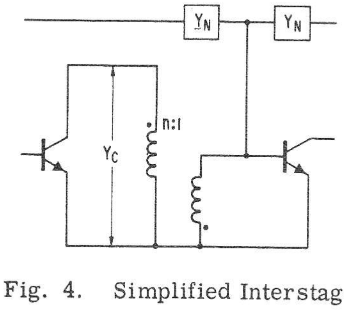

Fig. 4. Simplified Interstage.

It can be seen that compensation for the transistor conductance decrease is possible by arranging that the diode current increases as the emitter current decreases. However, each unit of conductance is accompanied by 0.2 units of susceptance so that perfect compensation is not possible. A choice therefore has to be made as to whether constant bandwidth or constant f_{max} is the most desirable feature; for the majority of applications constant f_{max} is the more useful and is chosen in this design. The use of shunting diodes has the additional advantage of providing an alternative path for the signal current thus reducing the emitter current change required for a given gain reduction.

CALCULATION OF TRANSFORMER WINDINGS

(1) Input Transformer

This must present a 50 ohm impedance to the source

$$n = \sqrt{\frac{G_s}{G_n + G_{11}}}$$

where

G_s = source conductance
G_n = neutralizing conductance
G_{11} = transistor input conductance ,

$$G_n = n_2 G_{12}$$

where

n_2 = ratio of interstages.

$$\text{Q of input transformer} = \frac{B_n + B_{11} + B_{strays}}{2\,(G_{11} + G_n)}$$

where

B_n = neutralizing susceptance
B_{11} = transistor input susceptance
B_{strays} = susceptance of strays

For $n_2 = 1.85$, $n_2G_{12} = 1.1$ mmhos, $G_s = 20$ mmhos, $G_{11} = 11$ mmhos, $\underline{n = 1.28}$, $nB_{12} = 2.5$ mmhos, $B_{strays} = 0.8$ mmhos, $B_{11} = 7$ mmhos. $Q = 0.43$, $L = -(B_n + B_{11} + B_{strays})$pf $= -10.3$ pf at 60 mc/s.

This Q is so low that its contribution to the overall bandwidth shrinkage may reasonably be neglected. There are then five remaining bandwidth determining tuned stages and if f_1, f_2, the lower and upper 3 db frequencies of the overall amplifier, are known the required interstage Q may be stated.

(2) Interstage Design

$$Q = \left[\frac{A^2 + B \pm B\sqrt{(A^2 + B)}}{2A^2}\right]^{1/2}$$

where

$$A = \frac{f_2^2 - f_1^2}{f_2^2 + f_1^2}$$

$$B = 2^{1/n} - 1$$

and where

f_2 = upper 3 db frequency of amplifier
f_1 = lower 3 db frequency of amplifier
n is number of tuned circuits (i.e. 5)

Inserting n = 5, f_1 = 50 mc/s, f_2 = 70 mc/s, $\underline{Q = 1.5}$

Coil should resonate at

$$f_o = Q\left[\frac{f_2^2 - f_1^2}{\sqrt{(4Q^2 - 1)(2^{1/n} - 1)}}\right]^{1/2} = 70 \text{ mc/s}$$

giving frequency of maximum gain

$$f_{max} = \sqrt{\frac{f_1^2 + f_2^2}{2}} = 61 \text{ mc/s}.$$

The turns ratio may be calculated from the following relationship derived in the appendix.

$$n = \sqrt{\frac{B_1}{B_2}}$$

where

$B_1 = B_{11}$ (imaginary part of y_{11}) + $\omega_o C_{\text{secondary strays}}$

$B_2 = B_{22}$ (imaginary part of y_{22}) + $\omega_o C_{\text{primary strays}}$.

Inserting

B_{11} = 7 mmhos
$\omega_o C_{\text{secondary strays}}$ = 1.8 mmhos
B_{22} = 1.35 mmhos
$\omega_o C_{\text{primary strays}}$ = 1.25 mmhos

$\underline{n = 1.85.}$

$$G_2 = \frac{2B_2 - \frac{2B_{12}}{n}}{Q} - \frac{G_{11}}{n^2} - \frac{2G_{12}}{n}$$

$$B_p = B_2 + \frac{B_1}{n^2} - \frac{2B_{12}}{n}$$

Primary = $\underline{-\ 6.7 \text{ mmhos}}$,

where

B_{12} = imaginary part of y_{12}
G_{12} = real part of y_{12}
G_{11} = real part of y_{11}
B_p = total susceptance referred to primary.

Inserting

B_{12} = - 1.35 mmhos
G_{12} = - 0.6 mmhos
G_{11} = 11 mmhos
Q = 1.5 mmhos,

G_2 = 3.95 mmhos.

But $G_2 = G_{22} + G_{damping}$. Inserting G_{22} = 0.5 mmhos,

$$G_{damping} = \underline{3.45 \text{ mmhos}}\ (310\ \Omega).$$

(3) Output Transformer

A turns ratio greater than unity was found to give spurious response due to leakage-reactance. Therefore $n = 1$ was chosen.

The secondary is center tapped to provide neutralizing, the tap down reducing the effect of the neutralizing susceptance on the transformer bandwidth.

$$B_c = B_{video} + \frac{Bn}{4} + B_2$$

where

$B_2 = B_{22} + B_{strays}$
B_c = total interstage susceptance referred to the primary
L_c = primary inductance
G_c = total interstage conductance referred to the primary,

and where

B_c = 3.1 + 0.5 + 2.5 = 6.1 mmhos
L_c = 0.35 μh

for $Q = 1.5$,

$$G_c = \frac{6.1}{1.5} = 4.1 \text{ mmhos}$$

$$G_{damping} + G_{video} = 3.6 \text{ mmhos } (280\Omega).$$

It was found, however, that a 200Ω damping resistor was required to give $Q = 1.5$, possibly because $K < 1$ and some regeneration from the earlier stages reflected a negative conductance across the output transformer.

A. G. C. CIRCUIT DESCRIPTION

In order to reduce the A. G. C. base voltage swing required, the emitter resistors of the three controlled stages are returned to -6V through a 6V zener diode connected to the -12V line.

The A. G. C. voltage is prevented from becoming positive by an FD 100 diode connected from the A. G. C. line to ground; this arrangement also gives an A. G. C. delay.

Stage voltage gain for $n = 1.85$ (no A. G. C.)

$$/A_V/ = \frac{-y_{21}}{n y_c}$$

where

y_{21} = forward transistor transadmittance

y_c = total admittance referred to primary of interstage

but from eq. (1A) in Appendix:

$$y_c (\omega_{max}) = G\sqrt{1 - \frac{1}{4Q^2}}$$

$Q = 1.5, \quad G = 7.2$ mmhos

$$/A_V/ = \frac{67}{1.85 + 7.2 \times 0.946} = 14.5 \text{ db}$$

Worst Case (Example 1):

Effect of A. G. C. on an Uncompensated Stage

$Q = 1.5$ i.e. $B_c = 5.5$ mmhos, $G_c = 3.7$ mmhos.

From curves of transistor y parameters versus emitter current.

At $I_e = 1$ mA (40 db A. G. C.)

$$\Delta G'_c = -1.95 \text{ mmhos}$$

$$\Delta B'_c = 0$$

due to y_{12},

$$\Delta y''_c = \frac{-y_{21}\, y_{12}}{y''_c}$$

where

$\Delta G'_c$ = change in conductance G_c due to transistor input and output conductance change with emitter current reduction.

$\Delta B'_c$ = change in conductance B_c due to transistor input and output susceptance change with emitter current reduction.

$\Delta G''_c$ = change in G_c due to transistor transadmittance change y_{12} and y_{21}.

$\Delta B''_c$ = change in B_c due to transistor transadmittance change.

$$\Delta G''_c = -0.28 \text{ mmhos}$$

$$\Delta B''_c = 0.105 \text{ mmhos}$$

$$\Delta G_c = -2.23 \text{ mmhos}$$

$$\Delta B_c = 0.105 \text{ mmhos}$$

$$\Delta G_c = \Delta G'_c + \Delta G''_c$$

$$\Delta B_c = \Delta B'_c + \Delta B''_c$$

Hence, new $Q = \dfrac{B_c + \Delta B_c}{G_c + \Delta G_c} = 3.7$,

$$\text{new } f_o = \frac{\omega_o}{2\pi}\sqrt{\frac{B_c}{B_c + \Delta B_c}} = 70\,(1 - .009) = 69.4 \text{ mc/s},$$

$$f_1 = \frac{\omega_o}{2\pi}\sqrt{1 - \frac{1 + \sqrt{4Q^2 - 1}}{2Q^2}} = 58 \text{ mc/s},$$

$$f_2 = \frac{\omega_o}{2}\sqrt{1 - \frac{1 - \sqrt{4Q^2 - 1}}{2Q^2}} = 77 \text{ mc/s}.$$

Stage BW = 19 mc/s. f_{max} = 68 mc/s.

Stage BW before A.G.C. = 82.5 - 23.6 = 58.9 mc/s.

f_{max} before A.G.C. = 61 mc/s.

From the above comparison the need of conductance compensation is quite clearly seen.

Worst Case (Example 2):

Effect of A.G.C. on a Compensated Stage

Assume same conductance and susceptance values as in Example 1.

Because of susceptance increase accompanying diode conductance increase one should under compensate the conductance decrease in order to restore the original f_{max}. Further as Q varies directly as susceptance and f_o only as the root of susceptance one is free to under compensate to some degree.

Let conductance obtained from diode = 2 mmhos, i.e. 0.25 mmhos of under-compensation. This is accompanied by $\frac{2}{5}$ = 0.4 mmhos susceptance, $Q' = \frac{B' + \Delta B_d}{G' + \Delta G_d} = 1.75$.

$$f'_o = 70\sqrt{\frac{B}{B' + B_d}} = \underline{67 \text{ mc/s}}$$

$$f'_{max} = f'_o\sqrt{1 - \frac{1}{2Q^2}} = \underline{61.25 \text{ mc/s}}$$

$$f'_1 = f'_o\sqrt{\left(\frac{f_{max}}{f_o}\right)^2 - \Delta} = \underline{35.6 \text{ mc/s}} : \quad \Delta = \frac{\sqrt{4Q^2 - 1}}{2Q^2}$$

$$f'_2 = f'_o\sqrt{\left(\frac{f_{max}}{f_o}\right)^2 + \Delta} = \underline{79 \text{ mc/s}}$$

$$BW' = \underline{43.4 \text{ mc/s}}$$

where

f'_o = new interstage resonant frequency due to susceptance change.

f'_{max} = new interstage frequency of max gain.

BW'= new bandwidth of interstage.

The overall f_{max} and BW change of the amplifier is of course less than that given above, and appears in Fig. 5 (a) and (b) for the uncompensated and compensated cases, respectively.

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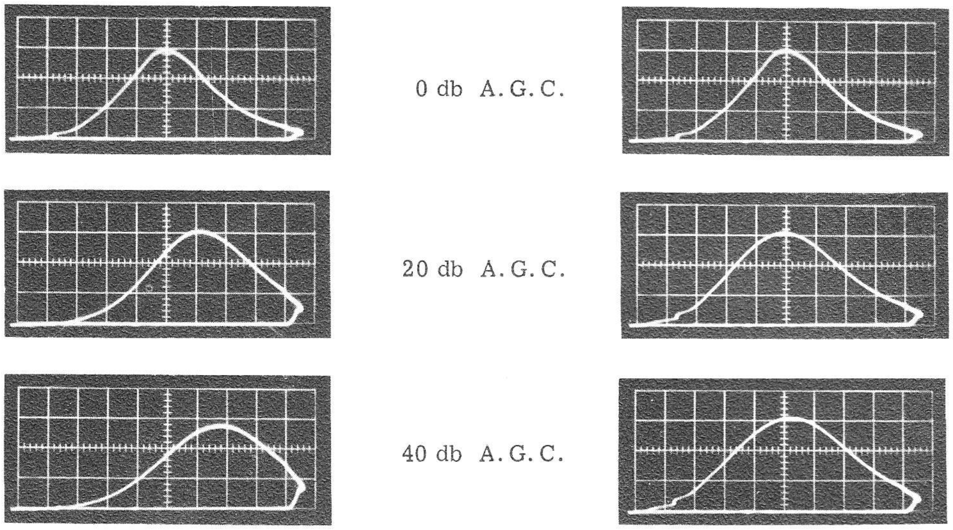
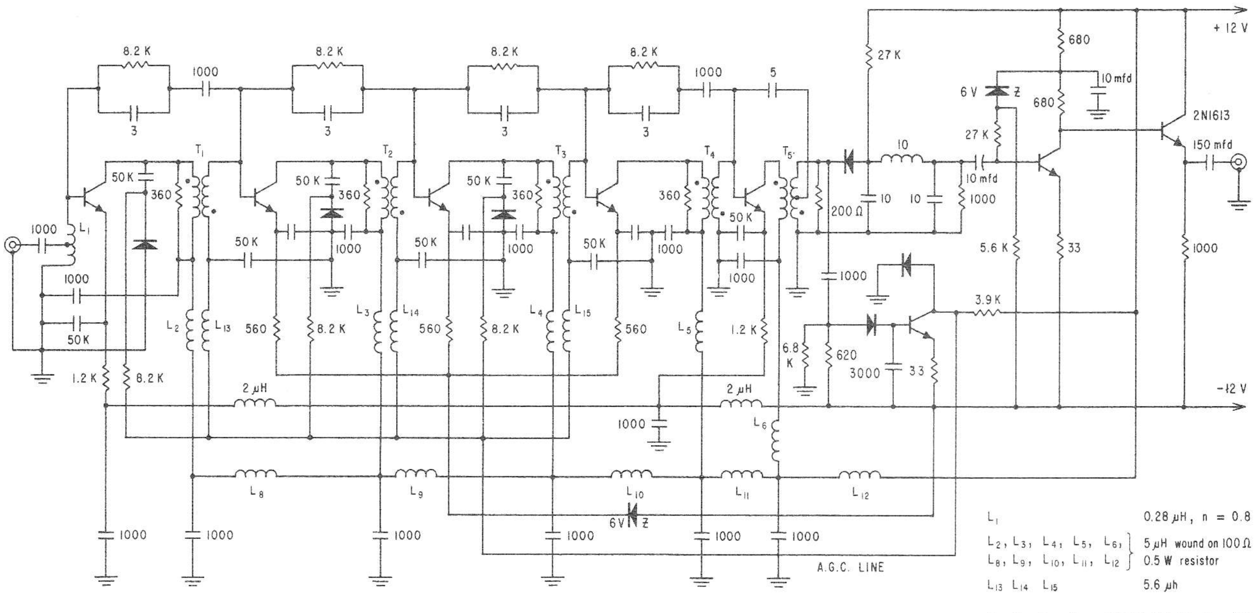


(a) Uncompensated Amplifier (b) Compensated Amplifier

Horizontally 5 mc/s per division

Note: BW appears less than 20 mc/s due to detector curvature

Fig. 5. Uncompensated and Compensated Amplifier Response for 3 Levels of A. G. C.



NOTE: All transistors 2N916 unless otherwise noted.
All diodes FD100 unless otherwise noted.
All capacitors in "pf" except where noted.

L_1	0.28 μH, n = 0.8
L_2, L_3, L_4, L_5, L_6, L_8, L_9, L_{10}, L_{11}, L_{12}	5 μH wound on 100 Ω 0.5 W resistor
L_{13} L_{14} L_{15}	5.6 μh
T_1, T_2, T_3, T_4,	Primary 0.4 μH, n = 1.85
T_5	Primary 0.35 μH, n = 1 Secondary center-tapped

$* \; n = \frac{\text{Primary turns}}{\text{Secondary turns}}$

Fig. 6. Schematic - 60 mc IF Amplifier.

APPENDIX

$$\text{Output voltage of one stage} = \frac{h_{21_e} i_{in}}{|y_o|}$$

where

$$|y_o| = \sqrt{G^2 + B^2}$$

$$= G\sqrt{1 + Q^2\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)^2}$$

for $\omega > \omega_\beta$, where $\omega_\beta = 2\pi f_\beta$, $\quad |h_{21_e}|(\omega) \simeq \frac{K}{\omega}$

where

$$f_\beta = \text{frequency at which } |h_{21_e}| = \frac{\beta_o}{\sqrt{2}}$$

$$V_{out}(\omega) \simeq \frac{K}{\omega G} \frac{1}{\sqrt{1 + Q^2\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)^2}}$$

Let $\frac{\omega}{\omega_o} = x$ where $\frac{\omega_o}{2\pi} = f_o$, the resonant frequency of the interstage

$$V_{out}(\omega) = \frac{K}{\omega_o G} \frac{1}{\sqrt{x^2 + Q^2(x^2 - 1)^2}}$$

$$\frac{dV_{out}(\omega)}{dx} = 0 \text{ for } \omega = \omega_{\text{max.output}}$$

Squaring the denominator and differentiating with respect to x,

$$d\frac{x^2 + Q^2(x^2 - 1)^2}{dx} = 2x + 4Q^2 x(x^2 - 1)$$

Set $2x + 4Q^2 x(x^2 - 1) = 0$

$$x_{max} = \sqrt{1 - \frac{1}{2Q^2}} \tag{1}$$

where

$$x_{max} = \frac{\omega_{max}}{\omega_o}$$

Bandwidth Shrinkage Due to Cascading n Stages

Consider a single stage at frequency $\frac{\omega}{2\pi}$ rads/sec.

$$V_{out}(\omega) = \frac{K_2}{\sqrt{x^2 + Q^2 (x^2 - 1)^2}}$$

where

$$K_2 = \frac{K}{\omega_o G}$$

for maximum output $x = x_{max} = \sqrt{1 - \frac{1}{2Q^2}}$

$$V_{out}(\omega_{max}) = \frac{K_2}{\sqrt{1 - \frac{1}{2Q^2} + Q^2 \left[(1 - \frac{1}{2Q^2}) - 1\right]^2}} = \frac{K_2}{\sqrt{1 - \frac{1}{4Q^2}}} \qquad (1a)$$

For overall 3 db frequencies i.e., response is $\frac{1}{\sqrt{2}}$ of that at $\omega = \omega_{max}$, and n stages

$$\frac{1}{\sqrt{2}} \left(\frac{K_2}{\sqrt{1 - \frac{1}{4Q^2}}} \right)^n = \left(\frac{K_2}{\sqrt{x^2 + Q^2 (x^2 - 1)}} \right)^n$$

Solving for $x^2 = 1 - \frac{1}{2Q^2} \pm \frac{\sqrt{(4Q^2 - 1)(2^{1/n} - 1)}}{2Q^2}$ (2)

For $Q \geqslant 2^{1/n - 1} \left[1 \pm \sqrt{1 - 2^{-1/n}} \right]$ equation (2) has two real roots, however, for $n > 1$ equation (1) limits $Q > 0.707$.

Select positive roots of equation (2) giving $\frac{\omega_1}{\omega_o}$ and $\frac{\omega_2}{\omega_o}$

$$x^2_{max} = 1 - \frac{1}{2Q^2} \quad \text{from equation (1)}$$

let

$$\frac{\sqrt{(4Q^2 - 1)(2^{1/n} - 1)}}{2Q^2} = \Delta$$

Therefore, $x_1^2 = x^2_{max} - \Delta$

$x_2^2 = x^2_{max} + \Delta$

i.e. the square of the normalized response has arithmetic symmetry.

$$x_2^2 - x_1^2 = 2\Delta$$

and

$$x_2^2 + x_1^2 = 2x^2_{max}$$

or

$$\frac{\omega_2^2 - \omega_1^2}{\omega_2^2 + \omega_1^2} = \frac{\sqrt{(4Q^2 - 1)(2^{1/n} - 1)}}{2Q^2 - 1} = A \qquad (3)$$

Let $2^{1/n} - 1 = B$.

Solving equation (3) for Q^2,

$$Q^2 = \frac{A^2 + B \pm \sqrt{B\,(A^2 + B)}}{2A^2}$$

This is the interstage Q required to give the 3 db frequencies $\frac{\omega_1}{2\pi}$ and $\frac{\omega_2}{2\pi}$.

$$\omega_o^2 = \frac{\omega_2^2 - \omega_1^2}{2\,\Delta}$$

$$\omega_{max}^2 = \frac{\omega_2^2 + \omega_1^2}{2}$$

where

$\frac{\omega_o}{2\pi}$ is resonant frequency of interstage

$\frac{\omega_{max}}{2\pi}$ is frequency of maximum gain.

Interstage Turns Ratio

$$Q = \frac{B_2 + 2n\,B_{12} + \frac{B_{11}}{n^2}}{G_2 + 2n\,G_{12} + \frac{G_{11}}{n^2}}$$

$$G_2 = \frac{B_2 + \frac{B_1}{n^2} - \frac{2B_{12}}{n}}{Q} - \frac{G_{11}}{n^2} - \frac{2G_{12}}{n}\,.$$

Total conductance referred to primary $= G_p$.

$$G_p = G_2 - \frac{2G_{12}}{n} + \frac{G_{11}}{n^2}\,.$$

Voltage gain at resonance $= \frac{-y_{21}}{G_p n}$.

Therefore, voltage gain maximum when nG_p is minimum:

$$\frac{d(nG_p)}{dn} = \frac{ndG_2}{dn} + G_2 - \frac{G_{11}}{n^2}$$

$$\frac{dG_2}{dn} = \frac{2B_{12}}{Qn^2} - \frac{2B_1}{Qn^3} + \frac{2G_{11}}{n^3} + \frac{2G_{12}}{n^2}$$

$$\frac{d(nG_p)}{dn} = \frac{2B_{12}}{Qn} - \frac{2B_1}{Qn^2} + \frac{2G_{11}}{n^2} + \frac{2G_{12}}{n} + \frac{B_2}{Q} + \frac{B_1}{Qn^2} - \frac{2B_{12}}{Qn} - \frac{G_{11}}{n^2} - \frac{2G_{12}}{n} - \frac{G_{11}}{n^2}\,.$$

Set $\frac{d(nG_p)}{dn} = 0.$

Therefore, $n\sqrt{\frac{B_1}{B_2}}$ and is independent of y_{12}.

A. G. C. Compensation

$$\omega^2_{max} = \omega^2_o \left(1 - \frac{1}{2Q^2}\right)$$

$$\omega'^2_o = \frac{B_c}{B'_c} \omega^2_o$$

$$\omega'^2_{max} = \frac{B_c}{B'_c} \omega^2_o \left(1 - \frac{1}{2Q'^2}\right)$$

where

$\frac{\omega'_o}{2\pi}$ is new resonant frequency for new susceptance.

Q' is new interstage Q.

For unchanged frequency of maximum gain,

$$\omega^2_{max} = \omega'^2_{max}$$

$$Q' = \frac{1}{\sqrt{\frac{B'_c}{B_c}\left(2 - \frac{1}{Q^2}\right) + 2}}$$

New Q should be Q' to compensate for $B'_c - B_c = \Delta B_c$ the susceptance change.

References:

(1) G. H. Booth, A High Frequency Transistor Amplifier Design Procedure. DRTE Report #1016, Department of National Defense, Canada.